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Doppler Radar in the 10 GHz Amateur Band Part-1

For years, a large selection of commercial intruder detectors has been available, based on the Doppler principle and operating in the micro-wave range. We wondered whether the simple 10 GHz transceiver with Gunn elements, which was so successfully used some years ago in the amateur radio world, could perhaps be used in Doppler radar equipment, similar to that used by the police to measure vehicle speeds on the roads.

As it turned out, we were able to obtain some surprisingly accurate readings with decidedly simple circuits! If developed a little further, the equipment could even be used to track aircraft and measure their distance and speed.

If you want to, you can, for example, monitor the speed of your neighbour's car electronically at any time! So now read on!

1. THE RADAR PRINCIPLE

Radar is an artificial word made up from the initial letters of "RAdio Detection And Ranging", which points immediately to the principle involved - determining the presence and the direction of a "target" with the assistance of radio waves and measuring how far away it is. For this purpose, the radar equipment's target must be irradiated with radio waves, which should be as strong as possible (Fig.1).

A small part of the high-frequency energy reaching the target is absorbed by it. The majority is scattered in many directions, and a small part is reflected back to the radar equipment. We know this from EME radio traffic (Fig.2).

If the radar equipment receives an echo from a specific object in the area on which

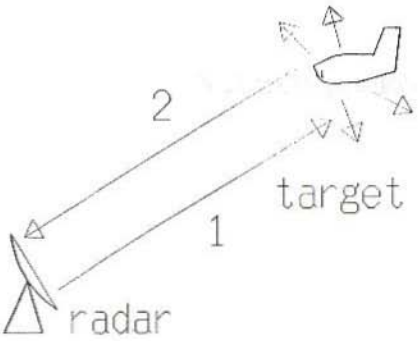

Fig.1: The Radar principle: the waves transmitted by the Radar set (1) are scattered in many directions by the target; a fraction is reflected back to the Radar

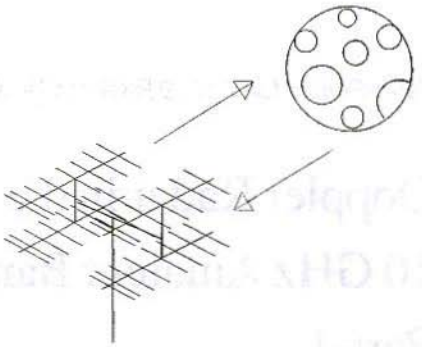

Fig.2: EME Mode, based on the same principle as Radar

it is trained, then there must be a target in this specific direction. This is called detection.

If we now measure the time elapsing between the transmission of the radar signal and the receiving of the echo, we can calculate the distance to the object - knowing, of course, that the radio waves are being propagated at the speed of light.

$$R = c * t/2$$

where:

R = Distance from radar to target
t = Time taken for signal to travel there and back
c = Speed of light (approx. $3*10^8$ m/s)

This is called ranging.

We can distinguish between two types of radar equipment:

→ Pulse radar
→ CW radar

1.1. Pulse radar

Pulse radar always transmits short pulses and listens to the echoes in between pulses. The pulse duration is typically in the approximate area of one per-thousandth of the reception time. For the descriptions in Fig.3, the so-called pulse-width repetition rate is thus:

$$X/(X+R) = 10\text{-}3.$$

The reception time is set in such a way that the transmission pulses have time to reach targets at the limit of the equipment's

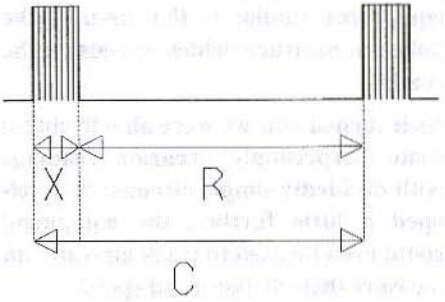

Fig.3: Pulse-width Repetition Rate: each cycle, C, consists of a Transmission Time, X, and a Reception Time, R

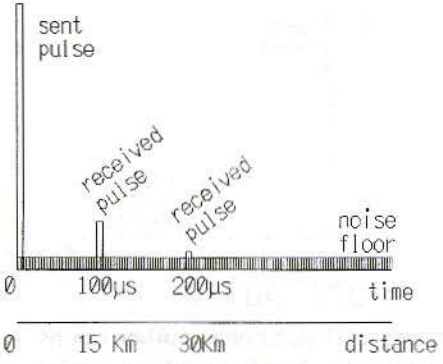

Fig.4: The echoes received from a Transmission Pulse

range and then return. A longer reception time would not make sense, as echoes coming in even later would be weaker than the receiver's limiting sensitivity. It is thus time to transmit a new pulse (Fig.4).

The pulse repetition frequency (PRF) is a very important parameter in radar technology (Fig.5). For the radar measurement of short distances, a high pulse repetition frequency is used because the echoes are already arriving shortly after the pulse has been transmitted. For large distances, on the contrary, the echoes need more time, so that a low PRF is indicated. In practise, the repetition frequencies lie between one and several kHz.

The briefer the pulse duration, the higher are the accuracy and the resolution of the distance measurement. Short pulses allow several targets close to one another to be differentiated (Fig.6a), whereas with longer pulses they appear as only one target (Fig.6b).

A magnetron is ideally suited to the generation of microwave pulses with high energy at a low pulse-width repetition rate. It has therefore been used for this since the 'forties, with a typical pulse-width repetition rate of about 1 part per thousand. The peak power can attain several megawatts.

But the transmission pulse duration can not be reduced indefinitely because, as the pulses become shorter, the reflected energy per pulse is reduced, and thus, of course, the energy of the echo. It must also be taken into account that shorter pulses have a greater signal band width.

1.2. CW radar

CW stands for continuous wave. As the name indicates, the transmitter is continuously switched on, but the frequency is switched between two or more values. The time between two frequency switchings must be sufficient for the signal to reach the target and return (Fig.7).

It has probably become clear that CW radar has nothing to do with the amateur radio meaning of "CW" as telegraphy mode.

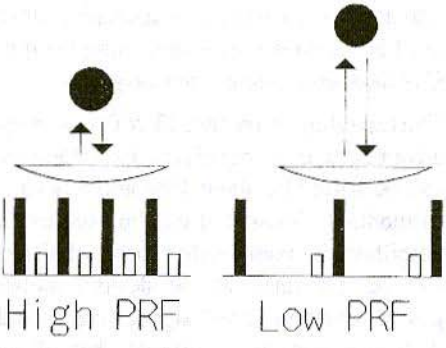

Fig.5: The Pulse Repeat Frequency must be matched to the distance to be measured: high for short range and low for long range

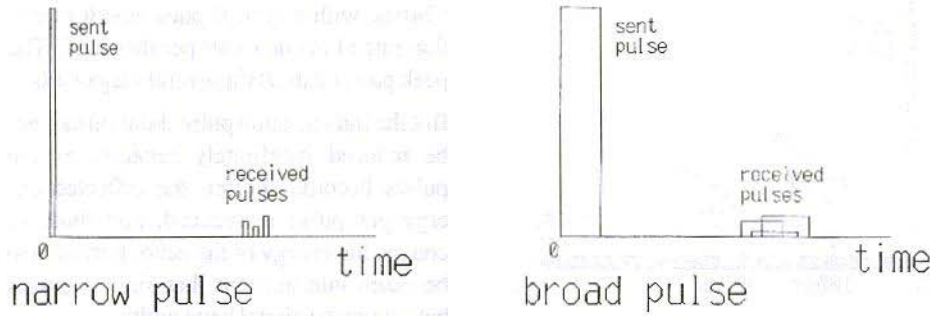

Fig.6: If the Transmission Pulse is short, targets close to one another can be differentiated. With long pulses, on the other hand, the echoes overlap

2. PRINCIPLE CONSIDERATIONS

The pulses from the classic pulse radar can be replaced by bursts on one frequency, with continuous wave transmission on another frequency in between. It is not all that simple to switch the frequency of a magnetron, except with special models. So several magnetrons must be used and must be switched at high power, which is also not a trivial matter. Although continuous magnetrons do exist (for example, those used in micro-wave ovens), using them in CW radar equipment is not easy either.

The travelling-wave tube (TWT) has many advantages in comparison. Its frequency can be varied by about 10% and it is easy to modulate because it can be used as an amplifier. So complicated forms of signal can be generated at the desired output power, even in small signal stages, and they are then much cleaner than if an attempt is made to modulate a magnetron accordingly (Fig.8).

With a travelling-wave tube, for example, coherent pulses can also be generated. Since the returning echoes are then also coherent, special filtering techniques can be used in the receiver which increase the signal-to-noise ratio or the range width. The disadvantage of the travelling-wave tube is that it can not generate the extremely high levels of power which characterise a magnetron.

For the radio amateur, of course, there is no question of using magnetrons, because in general they are much too powerful. One exception covers the types which are used in micro-wave ovens. They are cheaper than a 4CX250B, easily obtainable and also, if you buy a complete (old)

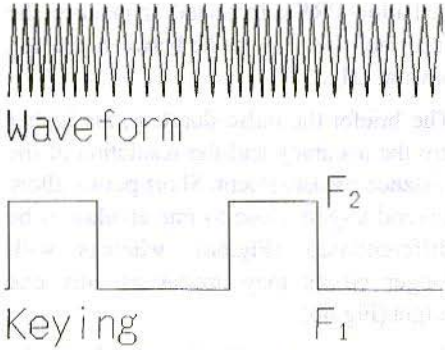

Fig.7: Waveforms of a Continuous Radar

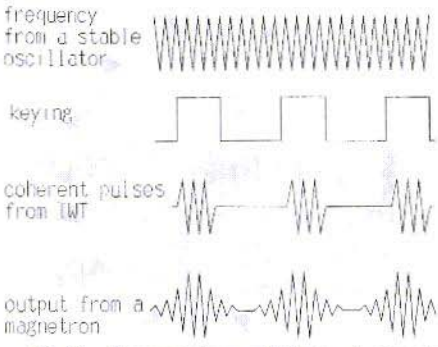

Fig.8: Comparison of Output signals from a Travelling Wave Tube and a Magnetron (only a few waves are shown out of the thousands in every pulse)

oven with them, equipped with a power supply and with cooling. With some comparatively simple modifications, the frequency can be moved to the 13cm

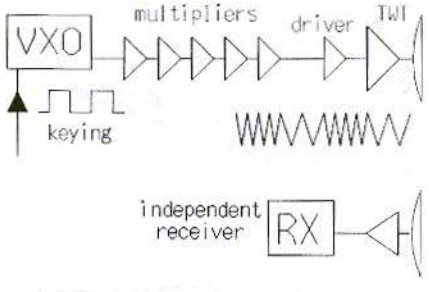

Fig.9a: Continuous Radar with FM

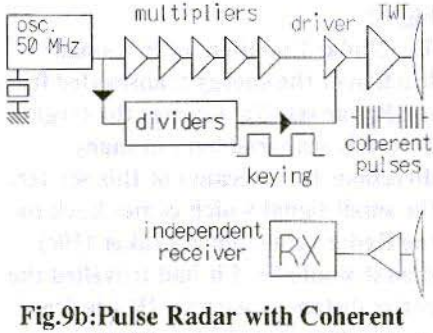

Fig.9b: Pulse Radar with Coherent PCM

amateur band and many hundreds of Watts of energy can be re-routed to an aerial. That would certainly be an interesting project (8). Switching the aerial between transmission and reception would, of course, not be exactly simple if you wanted to avoid baking the receiver input to a crisp.

Anyone who has the good fortune to get hold of a travelling wave tube amplifier can certainly also use it to carry out very promising experiments in the field of radar technology, as demonstrated by Fig's.9a and 9b.

For those amateurs who are really "poor" (i.e: lacking money, not spirit), the Gunn diode oscillator would seem to offer the only possibility for radar experiments. The obvious mode is CW, although we have read (1) that Gunn diodes can be forced into an intermittent high-powered mode if they are pulsed at an operating voltage which is higher than usual. But we did not investigate this type of misuse.

The lower microwave frequencies, such as the 23cm band, can be used for radar experiments, and normal ready-made FM transmitters probably can too.

The only problem would be switching the aerial between transmission and reception. You can certainly not switch a mechanical relay at a level of several kHz. So you would have to switch to a separate aerial for the receiver.

It should also be mentioned that the first radar sets operated in the VHF range, which could still be done today in principle. Even the short-wave range is used for radar, because there you can see "over the horizon" (remember the Woodpecker!).

2.1. The Radar Equation

We shall now carry out a theoretical study to determine the range width. The range width depends on the power received - more precisely, on the signal-to-noise ratio in the radar receiver.

We shall make use of the following values:

PE	Transmission power (W)
PR	Received power (W)
λ	Wavelength (m)
R	Distance from radar to target (m)
A	Reflected wave cross-section area of target (m^2)
G	Aerial gain

The power reflected back per solid angle unit is:

$$P_1 = \frac{P_E * G}{4\pi} \text{ (W)}$$

At the target, at a distance, R, from the radar, the power density is:

$$P_2 = \frac{P_E * G}{4\pi R^2} \text{ (W/m}^2\text{)}$$

The target reflects a power, P3, which is proportional to P2:

$$P_3 = A * P_2 \text{ (W)}$$

A, the radar back scatter cross-section, is a measure of the target's capability of reflecting radar waves. In warplanes, this value is made as small as possible.

P4 is thus the back-scattered power per solid angle unit:

$$P_4 = \frac{P_3}{4\pi} = \frac{A * P_2}{4\pi} = \frac{P_E * G * A}{(4\pi R)^2} \text{ (W)}$$

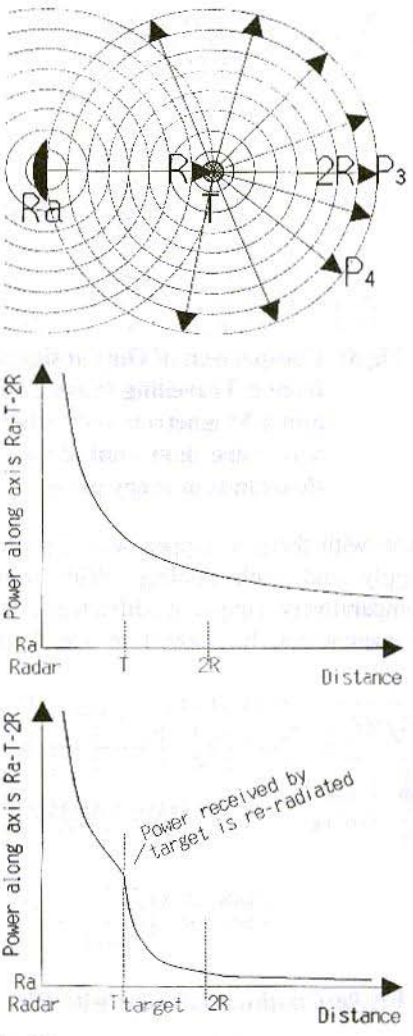

Fig.10:
The Radar Equation: only a small fraction of the energy transmitted from the Radar set, Ra, reaches the target, T, and is scattered here in many directions (a). Because of this scatter, the small signal which comes back to the Radar set is much weaker (10c) than it would be if it had travelled the entire distance - namely 2R - in free space (10b)

Finally, the following power density returns to the radar aerial:

$$P_S = \frac{P_E * G * A}{(4 \pi R)^2} * \frac{1}{R^2} = \frac{P_E * G * A}{(4 \pi)^2 * R^4} \ (W/m^2)$$

The aerial gain depends on its equivalent cross-section, Ae:

$$G = \frac{4 \pi A_e}{\lambda^2}$$

For a level wave front, the aerial therefore behaves like an absorbent aperture with a surface, Ae, of:

$$A_e = \frac{\lambda^2 * G}{4 \pi}$$

The signal supplied by the aerial to the receiver is thus P5 * Ae, which means that:

$$P_R = \frac{P_E * G * A}{(4 \pi)^2 R^4} * \frac{\lambda^2 * G}{4 \pi} = \frac{P_E G^2 A \lambda^2}{(4 \pi)^3 R^4}$$

If PR is the smallest signal usable for the radar set, then the maximum range width, R, can be read off from the following equation:

$$R^4 = \frac{P_E * G^2 * \lambda^2 * A}{(4 \pi)^3 * P_R}$$

The range width itself is thus the fourth power root of the fraction to the right of the equals sign. The problem in constructing radar sets for large range widths becomes clear here. The range width is proportional to the fourth power root obtained from the transmission power! So if, for example, you want to double the range width, then if everything else remains the same you must increase the power by a factor of sixteen!

We shall now try to understand the mathematics intuitively. As Fig.10a shows, the wave front is strongly curved at point Ra. As the distance grows, the radius of the curve becomes larger and larger, i.e. the wave front becomes flatter and flatter. If a wave front is strongly curved, the power diminishes very rapidly, which is shown by the steep sections of the curve in Fig.10b. If the wave front now becomes flatter and flatter, the power diminishes less and less, and thus the curve in Fig.10b becomes less and less steep.

At a great distance from the transmitter, the wave front is almost perfectly flat, so that the wave can cover great distances while losing almost no power. The power which transports the wave moves in only one direction.

Now what happens if an obstacle (radar target) is in the way? Right - the target scatters the waves in all directions again (Fig.10a), so that once again the propagation behaviour seen at point Ra returns. There is again a very steep fall in the (small reflected) power (Fig.10c). The power which strikes the target is proportional to ÖR. The power which reaches the radar receiver is thus proportional to $\sqrt{\sqrt{R}}$, or to $\sqrt[4]{R}$. In certain cases, this very high attenuation can be circumvented by building a transponder into the target. Whenever the transponder receives a signal from the radar, it amplifies it and loads it with information - such as identification and flying height - and then transmits the signal back to the radar set. The power received is then proportional to the fourth power root derived from R (Fig.11). It is clear that we can not make use of this possibility.

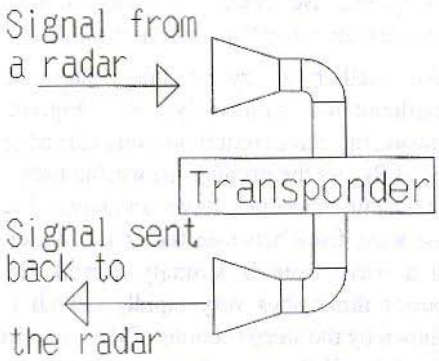

Fig.11: The Transponder Principle

The radar equation also shows that the 4th power also applies to the receiver sensitivity. The sensitivity has to be improved by a factor of 16 if the range width is to be doubled.

The smallest signal which the receiver can evaluate can also be expressed as the input noise power, N, for a given signal-to-noise ratio, SNR:

$$P_R = N * SNR$$

The radar equation then becomes:

$$R^4 = \frac{P_E G^2 \lambda^2 A}{(4\pi)^3 N * SNR}$$

The noise power, N, contains a fraction received from the aerial and a fraction generated in the receiver. The latter has many sources, among others the thermal noise. This can be very closely approximated by using the thermodynamic law:

$$N = k * T * B$$

where:

k is the Boltzmann constant: $1.374*10^{-23}$ W/Hz/K or Joule/K
T is the temperature of the object (K)
B is the receiver noise band width (Hz)

Now, we still haven't got the noise factor, n, which expresses the factor by which the receiver noise exceeds the minimum noise power established by the thermodynamic law. The total noise power is thus

$$N = n * k * T * B$$

Now we can write out the radar equation in a more comprehensive form:

$$R^4 = \frac{P_E G^2 \lambda^2 A}{(4\pi)^3 * SNR * nkTB}$$

The numerator still contains the reflected power, PE, and the received power, PR, has been resolved in the denominator.

As the final factor, we would now like to introduce the power losses between the generator and the aerial, as well as between the aerial and the receiver. The sum of these losses, L, comes into the denominator of the radar equation:

$$R^4 = \frac{P_E G^2 \lambda^2 A}{(4\pi)^3 * SNR * nkTB * L}$$

This should suffice for our purposes, and we shall now use this radar equation to estimate the range width of our experimental radar set-up (Fig.12). Let us assume the following values for our 10 GHz blow-through mixer with Gunn elements:

P =	$10mW = 10^{-2}W$
G =	20dB (or more) = $100 = 10^2$
	thus: $G^2 = 10^4$
λ =	$c/f = 3 * 10^8/10 * 10^9 = 30mm$
	thus: $\lambda^2 = 10^{-3}$
A =	$1m^2$
$(4\pi)^3$ =	$1.98 * 10^3$
SNR =	10dB = 10
n =	20dB = 100
	(not exactly state of the art!)
T =	$290 = 2.9 * 10^2$ K
B =	8 kHz = $8 * 10^3$ Hz

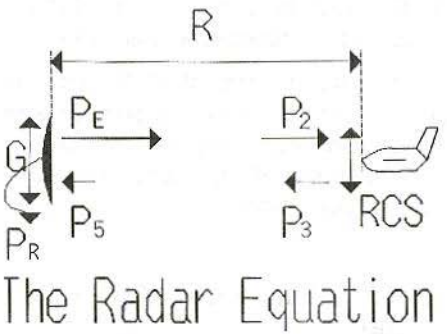

Fig.12: The most decisive values from the Radar Equation

This is an adequate band width, for if the Doppler frequency reaches 8 kHz, that corresponds to a target speed of 459km/h.

$$L = 6dB = 4$$

We now introduce this value into the above radar equation and we get: R = 140m.

That might appear somewhat optimistic for the minuscule transmission power, but we shall find out by practical experiments. Initially we want to try and increase our range width on paper. If we had a hypothetical receiver without noise and without losses, only the thermal noise would remain and, with SNR = 1, we would obtain n = 1 and L = 1 from our radar equation, if the other parameters remained unaltered, giving R as 1.12km!

With a modern receiver and small power losses, the range width will thus lie somewhere between the extremes 140m and 1km.

We can now lay the receiver aside and juggle with the other parameters instead:

a) Increasing Aerial Gain

With a reflector diameter of 160cm, we would obtain a gain of about 40 dB, i.e. a hundred times the earlier figure, so that the range width is multiplied by ten: R = 1400m. Disadvantage: targeting with a large aerial is very much more difficult.

b) Using lower frequencies

This makes the wavelength bigger - for example about 10 times as big if we use the 23cm band. The range width is increased by the square root of 10 = 3.16. Thus R = 440m. Disadvantage: as we are still calculating on 20dB gain, the aerial becomes correspondingly larger!

c) The last possibility - increasing the transmitter power

Unfortunately increasing the transmission power by ten only increases the range width by the fourth power root of 10, which is 1.77. So a range width of 140m becomes 249m at a transmission power of 100mW, or 442m at 1W, or 787m for 10W, and so on.

A super 3cm radar set with a transmission power of 10W from a travelling-wave tube and a 1.6m parabolic antenna could have a range width of between 7km and 30km, depending on the quality of the receiver.

A 23cm radar set with a transmission power of 100W and an antenna gain of

20dB (super long Yagi) would have a range width of between 4km and 30km.

So we can see that it is theoretically possible to track aircraft - especially large aircraft - and measure their speed. Now it's time to deal with the radar backscatter cross-section (RCS).

2.2. Radar Backscatter Cross-section (RCS)

The higher the RCS value, the greater the range width. All range widths given above are based on an RCS value of $1m^2$. It is clear that, for example, a Boieng 747 has a much higher RCS value. Moreover, it is immediately clear that the RCS value is always variable, depending on the orientation of the target relative to the wave front of the radar signal (Fig.13).

The RCS value is also dependent on the shape of the target. Flat surfaces act like mirrors and reflect the waves very well. Sharp edges, slots and points can act as aerials and reflect the microwave energy received. Multiple reflections take place in cavities or intersecting surfaces, and the incoming wave can be reflected to the radar set (Fig.14).

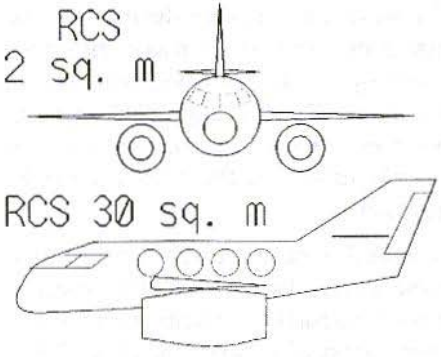

Fig.13: The Radar Backscatter cross-section, RCS, depends, among other things, on the orientation of the target

But the RCS value is not dependent on the size and shape of the target alone. The material it's made of also has an influence. Metal reflects much better than, for example, plastic or composite materials. Certain materials which absorb microwaves are used to reduce the RSC value of warplanes, as is special shaping (remember the stealth bomber). Some aircraft optimised in this way should allegedly have an RCS value no greater than that of a seagull!

If we now target a large (civilian) aircraft with an RCS value of $50m^2$, the theoretical range width of our radar will be multiplied by the fourth power root of 50 = 2.6. So the original range width of 140m becomes about 400m.

Now here are a few examples of RCS values:

Human being:	0.1 - 1.9 m^2
Seagull:	0.01 m^2
Fly:	0.00001 m^2
Aircraft:	0.5 m^2 (head-on)
	20 m^2 (sideways-on)

A more precise RCS value can be calculated only for simple surfaces (flat surfaces, spherical surfaces and the like). A multiform object can be seen as a combination of many simple shapes. Some of them will be good reflectors, others bad. To the radar set, such objects appear as a collection of bright spots between dark spots.

Depending on whether the waves from the various reflecting part-surfaces finally reach the radar receiver more in phase or more in opposite phase, the target is bright or dark, the RCS value large or small. If any of the structures of the target at all is in

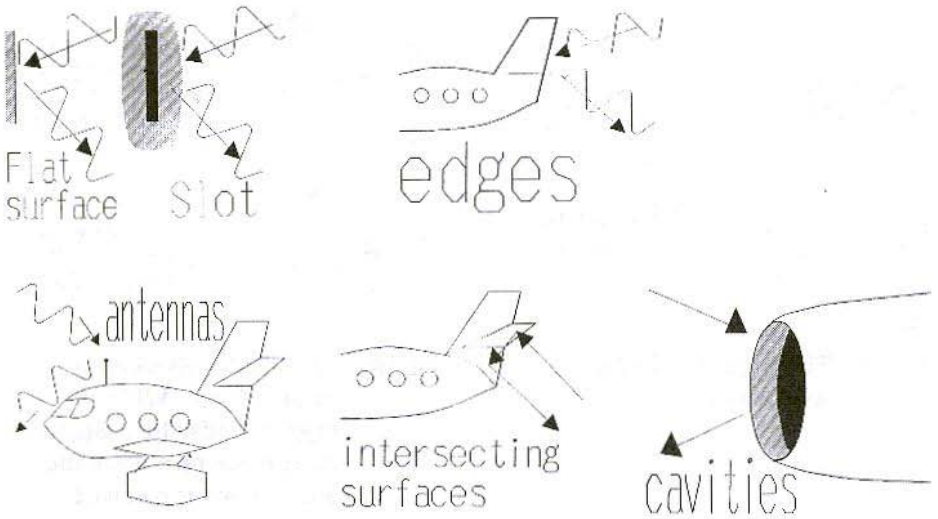

Fig.14: Certain structures of a Radar target can reflect particularly well, especially if their dimensions are of the order of magnitude of the wavelengths used; this creates particular;y bright spots in the Radar image

resonance with the radar wavelength, this produces a particularly bright spot, because the reflection is particularly effective.

It should now no longer come as a surprise to find that the RCS value of a moving target is continually varying in practise, because the phases of the wave fractions reflected are continually being superimposed on each other. Variations in the propagation conditions in the atmosphere, which are expressed as fading with changing strength and time constants, also play a part. Each individual radar echo will thus have a different intensity. There will be an average value, with a random scattering around it.

Thus, when signals are processed in the receiver of a pulse radar, a number of successive echoes must be determined in order to increase the accuracy of the process.

2.3. The Doppler effect

The Doppler effect is familiar to everyone since rail and road traffic became widespread. This effect has become very important in astronomy (Fig.15), as the so-called red shift of spectra, on the basis of which their speeds and distances can be calculated on the cosmic scale.

The Doppler effect affects all electromagnetic waves, and this includes microwaves. Here it can even be measured quite handily. The frequency shifted by the Doppler effect can be calculated using the following formula - with the proviso, of course, that the speed of the object is small in comparison with the speed of light:

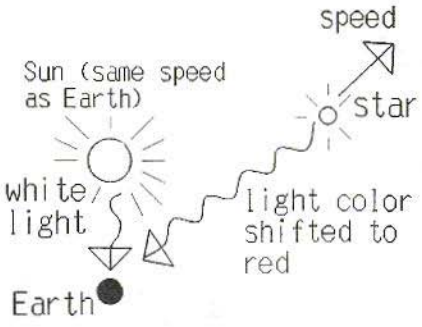

Fig.15: The Doppler effect in Astronomy

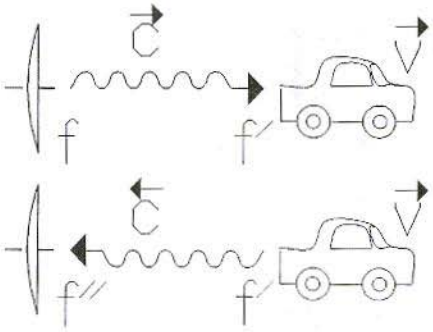

Fig.16: The Doppler effect in its dreaded form. With Doppler Radar the Doppler Effect is doubled when the return signal is received

$f' = f * (1 - (v/c))$

f'= the frequency received
f = the frequency transmitted
v = speed of object
c = speed of light

But we must remember that this formula is based on the assumption that the frequency, f, is generated by the moving object itself, not by the observer.

2.4. Doppler radar

In a Doppler radar system, the output frequency, f, is generated, not by the moving object, but by the radar set. When it encounters the moving object, a Doppler frequency shift has already taken place. This displaced frequency, f', is reflected back to the radar set and undergoes a second Doppler shift. The reception frequency, f'', thus includes a double frequency shift (Fig.16).

$f'' = f * (1 - (2\ v/c))$

The Doppler frequency, fd, is the difference between f and f'', i.e. the total value of the frequency shift which the original frequency, f, has undergone due to the Doppler effect.

$f_d = f * 2\ v/c$ oder $v = f_d * c/2f$

As we can see from this formula, the speeds of the target, v, and the Doppler frequency, fd, are directly proportional to

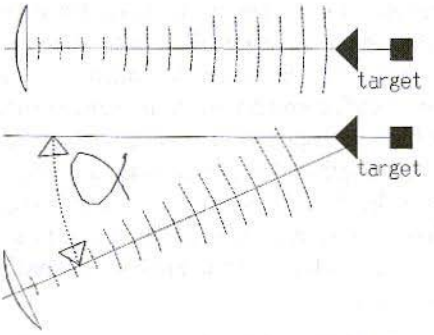

Fig.17: The Alpha Angle must be taken into account in the calculation of the speed

Doppler Frequency Hz	Speed m/s	Speed km/h
0	0	0
10	0.146	0.526
50	0.731	2.633
100	1.463	5.266
200	2.956	10.532
500	7.315	26.33
1000	14.63	52.66
2000	29.26	105.32
5000	73.15	263.3

Table.1: Doppler Frequencies and Speeds for a Transmission Frequency of 10.25 GHz

one another. We can also see why it is particularly easy to deal with the Doppler effect in the micro-wave range. For speeds which arise in practise, Doppler frequencies are in the low- frequency range. able 1 gives some examples of this, based on the assumption that the transmission frequency of our radar set is 10.25 GHz.

As we see, an interesting speed range can be covered using a handy receiver range width. The Doppler formula was averaged to obtain the figures for the table. Multiplying the frequency by $1.463*10^{-2}$ gives the speed in m/s. The factor for the speed in km/h is $5.266*10^{-2}$.

So far we have been assuming that the target is moving directly towards or away from the radar set. But in practise this is extremely rare. There is almost always an angle, alpha, between the direction of propagation of the radar waves and the direction of motion of the target (Fig.17).

The Doppler equation must therefore be expanded as follows:

$$f_d = \frac{f * 2v}{c} * \cos\alpha$$

or

$$v = \frac{f_d * c}{2f * \cos\alpha}$$

If the radar set is standing at the side of the road and the target is picked up when it is a few tens of metres away, cos alpha is so close to 1 that the term can be ignored for practical purposes. If the alpha angle is 33.5°, cos alpha = 0.83. The speeds listed in Table 1 would need to be multiplied by 1.2 in such a case.

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Doppler Radar in the 10 GHz Amateur Band Part-2

3. SELECTION OF AN SHF RADAR SETUP

Several choices are open, offering varying complexity and results.

3.1 The "Gunnplexer"

This device, once widely used, is based upon a circulator (Fig.18). Power derived from a Gunn oscillator reaches the antenna via a circulator, with a small component reaching the diode mixer as a heterodyne signal. Received signals reach the mixer practically unhindered by the circulator. The desired output signal is the difference between the received and the local oscillator (LO) signal.

In this case the less than perfect directionality of the circulator is a desirable feature, since the LO works as a transmitter and at the same time makes available a heterodyne frequency. The reverse attenuation of the circulator is around 20dB so that even with several hundred mW transmit power the mixer diode is not threatened. In addition the circulator protects the LO from the outside world, which is good for stability.

A frequency drift of, for example, 10 MHz at 10 GHz is just one part in a thousand; this also means that any error in the Doppler frequency is also only 1 in 1000. There are no other errors, since the transmit and heterodyne frequencies are produced by the same oscillator.

With a 40cm dish antenna we achieved ranges of 200 metres.

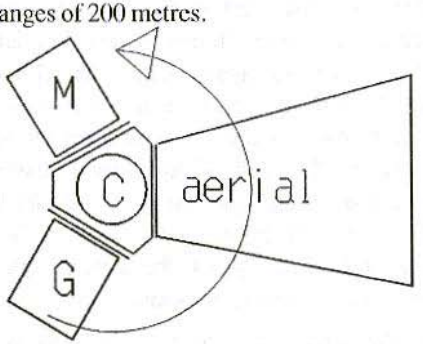

Fig.18: Practical possibility using a Circulator (i.e: Gunnplexer)

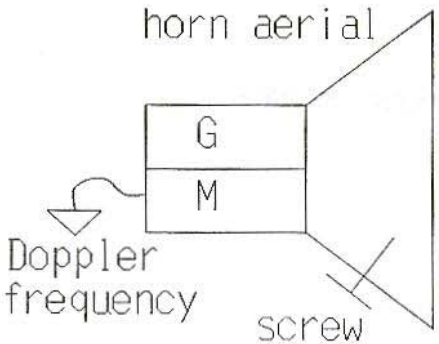

Fig.19: Another alternative: a Burglar Alarm. The minute antenna rules out large radar ranges

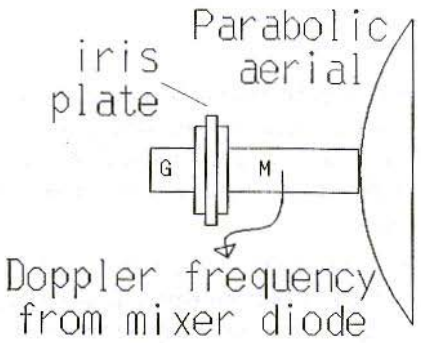

Fig.20: A 'Blast-through' Mixer for broadband FM is a good choice for Radar experiments

3.2 Burglar alarm

The setup sketched in Fig.19 was used some years back in microwave burglar alarms. Use was again made of the Gunn oscillator in the upper module, the transmitter power feeding a horn antenna. A screw in the wall of the horn radiator reflected a small part of the transmitted power into the lower module (the receive mixer). Regrettably half the receive capability was lost in the transmitter.

At the same time, the lack of isolation meant that nearby environmental influences could cause significant drifting, especially with movement directly in front of the antenna (which was what the setup was designed for). In our case the high level of drift led to inaccurate Radar measurements. Another disadvantage is that this assembly cannot be attached to standard waveguide, which makes application with a dish antenna problematic.

Using the horn antenna supplied, very short ranges are achieved and then without great accuracy.

3.3 The "blast-through" mixer

This arrangement was very popular in amateur radio circles some years back since it is easy to make at home and gives good results. A Gunn oscillator is separated from the rest of the assembly by an iris (a plate with a small round hole). According to the diameter of the hole and the thickness of the sheet, a greater or smaller part of the oscillator output reaches the antenna through the mixer (Fig.20). This (low) power is the output signal and at the same time the heterodyne signal.

Being a non-linear element, the mixer diode produces the Doppler signal - an audio frequency signal equal to the difference between the transmit and receive frequencies.

The method of operation is certainly very similar to the Gunnplexer but the transmit output is significantly lower; a few mW instead of 40mW! With an iris of 6mm to 7mm diameter made of 1mm brass sheet and a Gunn oscillator, the approx. 5mW produced is not optimal but adequate for the mixer diode. If we make the iris larger we certainly get more mixer current flowing but the oscillator also loses stability. With high powered Gunn oscillators the mixer diode can even be damaged.

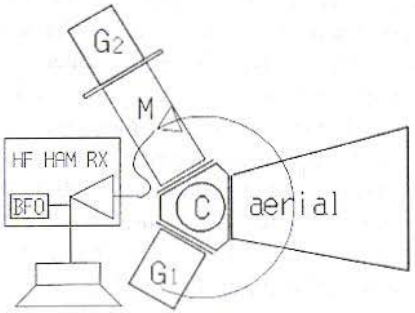

Fig.21: An attempt to increase range by using an IF of 30 MHz was abandoned due to frequency unsuitability

The transmit power will not exceed a few mW with a blast-through mixer. In the past we managed with this amount in wideband FM to span 130km and more, from one Swiss mountain top to another. Yet the range for Doppler Radar is no more than 100 to 150m, as our practical tests demonstrated.

3.4 Higher IFs

In the setups described so far the incoming signal has been converted in the mixer diode down to the audio frequency range.

Down there is also where the strongest phase noise effects of the oscillator appear, and Gunn oscillators are notorious for phase noise. The phase noise intrudes on the audio signal and reduces the sensitivity of the system.

One probable improvement would be to use a GaAsFET oscillator stabilised with a dielectric resonator (5) giving lower phase noise, whilst on the other hand it would be interesting if introducing an IF would increase sensitivity.

In our FM transceivers we used an IF of 30 MHz. So we experimented with the setup shown in Fig.21. The first oscillator G1 produces the transmit frequency f at the highest possible level, the second oscillator being adjusted to f + 30 MHz or f - 30 MHz.

The signal reflected by the target lies between f + 8 kHz and f - 8 kHz, depending whether it is moving towards or away from us, and at what speed. Behind the mixer the signal therefore lies between 29.992 and 30.008 MHz. In the receiver it is amplified several score times and, if we switch in the BFO, we get an audio frequency tone in the loudspeaker. We can adjust the BFO so that this tone amounts to a few Hz when the target is receding at maximum speed from the Radar device; when the target is static the audio frequency is 8 kHz, and when it is approaching fastest the tone is 16 kHz (provided the receiver bandwidth can handle this).

If the target's direction of movement is not of interest, we can adjust the BFO so that the tone is around zero when the target is static.

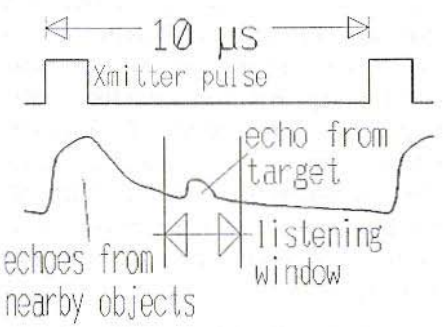

Fig.22: A Pulsed Transmitter and a brief time window can avoid receiver overload and clutter arising from reflection from nearby objects

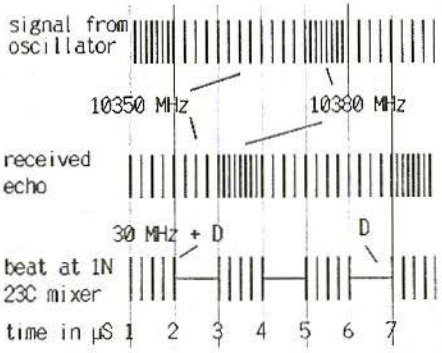

Fig.23: Frequencies in a Frequency-Inverted Continuous Wave Radar based on the 'Blast-through' Mixer. A low frequency is produced in the Mixer Diode between the received pulses: this is the Doppler Frequency

When it now moves, the audio frequency will rise in either case up to a maximum of around just 8 kHz - which corresponds already to a speed of about 450km/h.

The sensitivity of this setup should exceed that of those described previously by a wide margin and to give an example: a bird that flew away a few score metres already gave a very strong signal! We calculate that the range with this setup for a target with a Radar cross section (RCS) of one square metre is several hundred metres. Yet, as we know from our endeavours with the Radar formula, the range can barely exceed 1km - even if we had a noise-free receiver.

This setup using an IF has, however, two serious deficiencies.

Firstly there is breakthrough from the power oscillator G1 through the circulator into the receiver and, small though it is, it exceeds the receive power by several orders of magnitude. This leads to what we have come to know in VHF contests - desense. The idea of using a totally separate transmitter with its own antenna does not help either because reflections from nearby objects and/or side-lobes of the antenna will still cause sufficient coupling of transmit power back into the receiver that blocking effects arise.

Secondly there is the thermal drift of the Gunn oscillator. The calculation of the Doppler frequency depends on a precise 30 MHz difference between transmitter and receiver. This, however, is difficult if not impossible to maintain. Just breathing on one of these oscillators will cause a frequency change of a couple of kHz!

It is possible that the introduction of dielectric resonators (5) and/or PLL circuits could bring the required stability. If only we could get the two oscillators to drift at the same rate, the Doppler error could be reduced to negligible proportions, as we saw in the Gunnplexer.

3.5 Blanking the transmitter using window techniques

A further possibility for improving things lies in keying the transmitter on and off and operating the receiver when we expect an echo from the target (Fig.22). The receiver would then maintain a defined window. During this window period the transmitter would be turned off and the near-target echoes would already be dying away.

How will the Gunn oscillator react to having its power supply keyed on and off? The frequency will presumably have to stabilise each time after switching on and

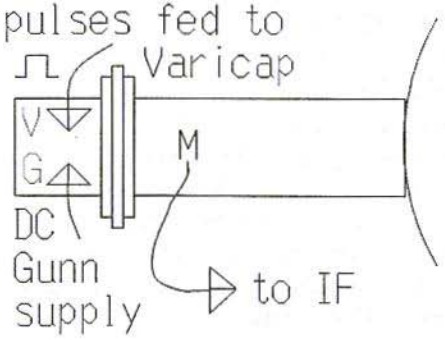

Fig.24a: Another way of avoiding Rx overload by the Tx pulse is Periodic Frequency Inversion, which would allow an IF of 30 MHz, which wouldincrease sensitivity

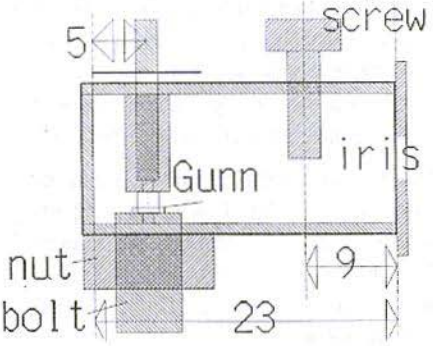

Fig.24b: Side view of a Frequency modulated Gunn Oscillator Dimensions in mm.

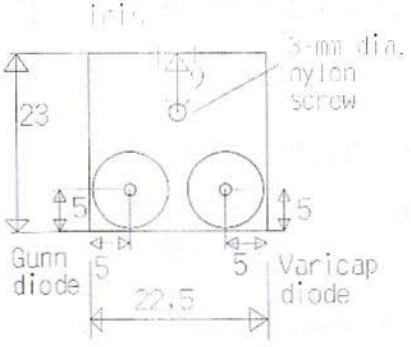

Fig.24c: Top view of the same oscillator as in Fig.24b

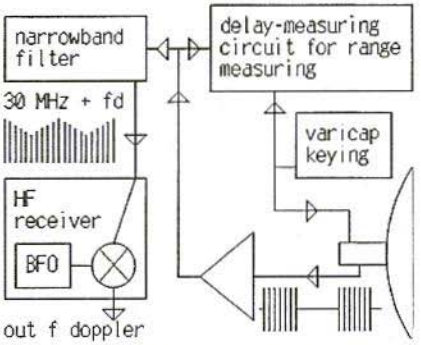

Fig.25: Block Diagram of a CW Radar with Frequency Inversion based on the 'Blast-through' Mixer

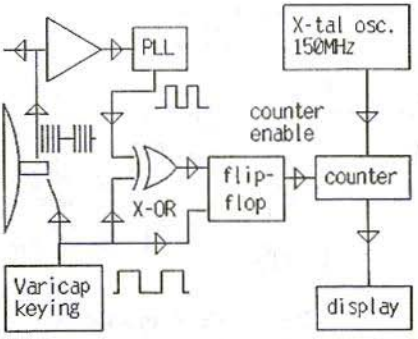

Fig.26: Distance Measurement: the Counter is started with the first Pulsed Transmitted and stopped at the beginning of the Received Pulse

would not be constant throughout the pulse. However, this will not do for Doppler measurements.

A PLL will not lock in within the space of a couple of microseconds.

If we achieve the keying of the transmitter with a PIN diode its reverse attenuation becomes a critical factor. If we attempt mechanical keying, for example with a

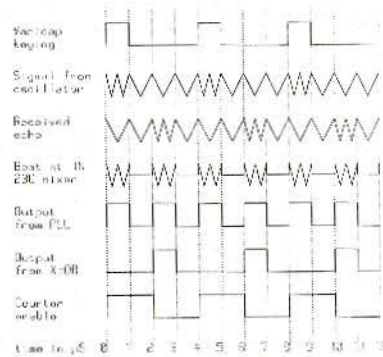

Fig.27: Timing Diagram from Fig.26

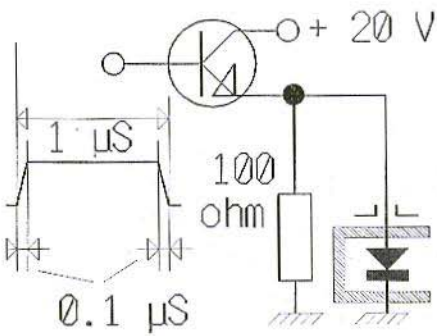

Fig.28: Pulses for the Capacity-Diode must have steep edges to enable the Microwave Frequency to be switched between the two values as rapidly as possible

propeller, the rapidly rotating rotor blades would bring about some "interesting" doppler effects.

3.6 Frequency keying

A better scheme would doubtless be to leave the transmitter G1 running all the time, but shift its frequency. It then transmits only short pulses with the frequency f and for the rest of the cycle is detuned by several score MHz by means of a varicap diode. This also avoids overloading the receiver while listening to echoes (Fig.23).

We carried these thoughts across to the blast-through mixer by equipping it with a varicap diode (Fig's.24a, b and c) supplied with keyed direct current. After this we are still left with no more than an oscillator and the problems of differential drift of the two oscillators remain.

With the knowledge that c = 300m in 1 microsecond, we know that our signal needs 2 microseconds to reach and return from a target 300m away. So in the time interval t we will switch a 20V impulse of 1 microsecond duration onto the varicap diode of the Gunn oscillator. This shifts the frequency of the oscillator by, for example, 30 MHz upwards, for one microsecond, then it reverts to its original frequency. We assume a mark-space ratio of 25 per cent.

Thus, we see on the IF side of the receive mixer a series of bursts with the frequency 30 MHz + doppler frequency. The period of the burst takes 2 microseconds which corresponds to a pulse repetition frequency (PRF) of 500 kHz.

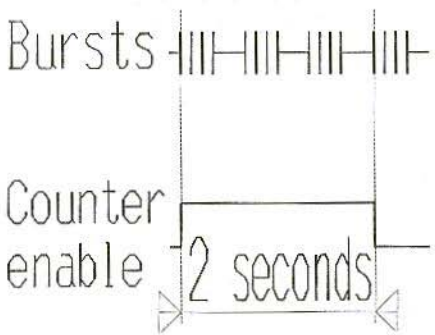

Fig.29: To increase accuracy the Counter can be enabled for several bursts

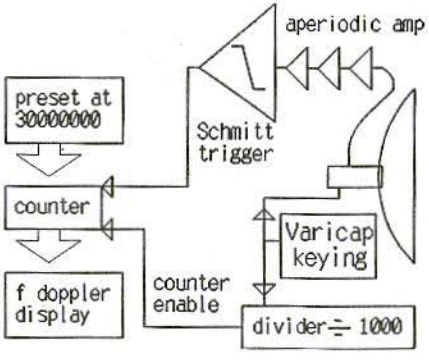

Fig.30: A Counter replaces the HF Receiver for speed measurement

This signal is now amplified in a low-noise 30 MHz amplifier, passed through a narrow-band 30 MHz filter and cleaned up in this way, is passed to the HF receiver. It is still 30 MHz + the doppler frequency with a residual 500 kHz AM, whose level depends on reflections in the filter and its bandwidth. After mixing with the BFO signal we are left for evaluation an audio frequency that is exactly the doppler frequency. Fig.25 shows this scheme schematically.

3.7 Concepts for measuring distance gand speed

With the setup described above distance measurement can be carried out by evaluating the time between the pulse to the varicap diode and the reception of echoes. The principle is shown in Fig.26.

With the leading edge of the pulse that keys the varicap diode a counter is started; it is stopped again by the leading edge of the pulse from the EXOR gate (Fig.27).

With the knowledge that the distance r = t *c/2 it can be calculated that a measured time period of, say, 2 microseconds implies a distance of 300m. To get a figure-perfect display we let the counter

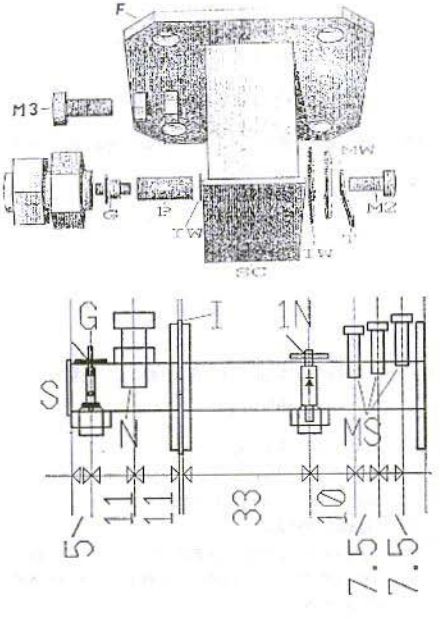

Fig.31a: The Gunn Oscillator in 3-D
Fig.31b: Dimensions of the Oscillator and Mixer

F = Flange
M3 = Metal or Nylon Screw, matching the bearing nuts
G = Gunn Element with M5 nut and bolt for adjustment
P = Metal bolt (copper or brass)
IW = Insulating Washer, inner and outer (PTFE)
SC = Waveguide Short-Circuit
T = Solder Tag
M2 = Steel Screw, M2
MW= Metal Washer
S = Short Circuit (on the plate soldered on the end of the waveguide
N = Nylon Screw, 3 to 6mm in diameter
I = Iris Plate (brass, 1mm thick with 6mm hole in the centre)
1N = 1N23 Mixer Diode
MS = Matching Screws (brass, 3mm diam)

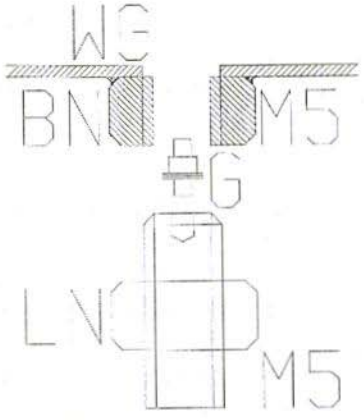

Fig.31c: Mounting details of the Gunn Element

WG= Waveguide (inside)
BN = Bearing Nut (M5, steel) soldered to the Waveguide
G = Gunn Element, larger flange grounded
LN = Nut (M5, steel)
M5 = Support Peg (M5, steel) bored axially for the end of the Gunn Element

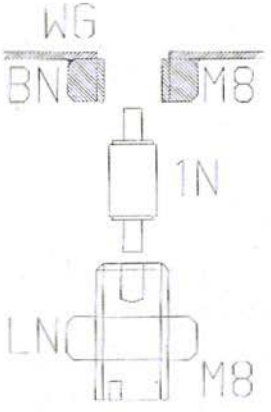

Fig.31e: Detail of the Mixer Diode inset

WG= Inside of Waveguide
BN = Bearing Nut (M8, steel) soldered to the Waveguide
1N = Mixer Diode
LN = Nut (M8,steel)
M8 = Support Peg (M8, steel) bored centrally for the Diode

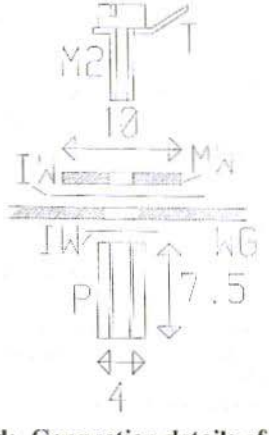

Fig.31d: Connection details of the Gunn Element

M2 = M2 steel Screw
T = Solder tag
MW= Metal Washer, 10mm outside diameter
IW = Inner/outer PTFE Insulating washers
WG= Inside of Waveguide
P = Round peg of brass or copper. Length 7.5mm, outer diameter 4mm, bored axially and tapped M2

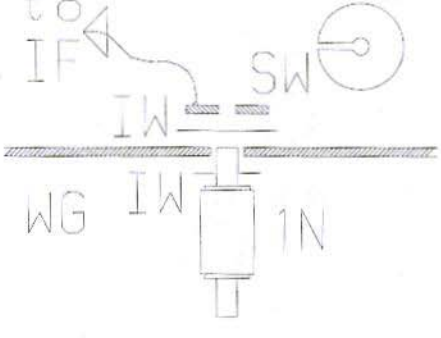

Fig.31f: Detail of the Mixer Diode connections

WG= Inside of Waveguide
IW = Insulating Washers (inner and outer PTFE)
1N = Mixer Diode
SW = Special Washer, fastened to the Diode connection and connected to the AF or IF preamplifier

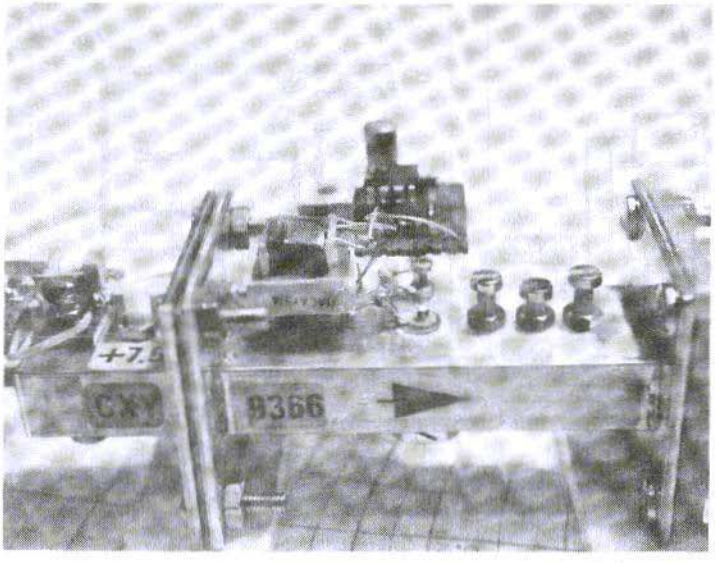

Fig.31g: Our SHF unit

count the pulses of a 150 MHz crystal oscillator: in 2 microseconds we get 300 pulses meaning that we can calibrate the counter display directly in metres.

If everything functioned perfectly - which of course it won't do - we could use this

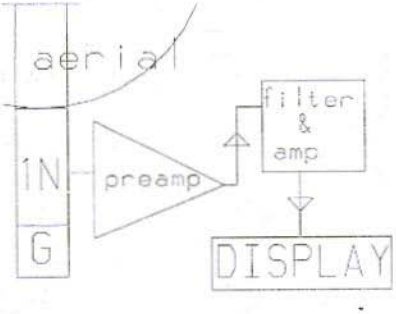

Fig.32a: Block principles of Doppler Radar system

Aerial =IKEA Dish with Penny Feed
1N = Mixer, connected to Antenna and oscillator by Waveguide
G = Gunn Diode
Preamp = see Fig.32b
Display = see Fig.35

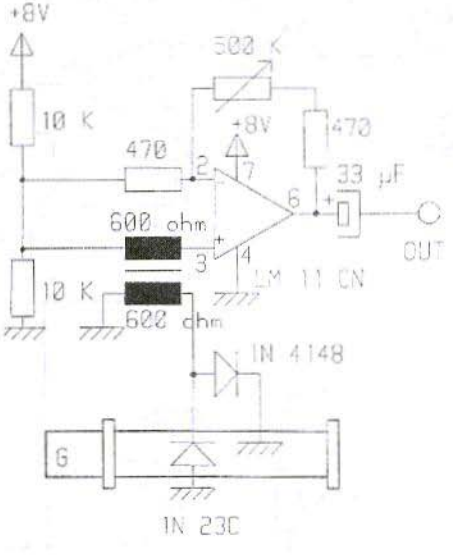

Fig.32b: Circuit of the Audio Frequency Preamplifier. The 1N23 is very sensitive to static electricity and is protected by the Switching Diode.

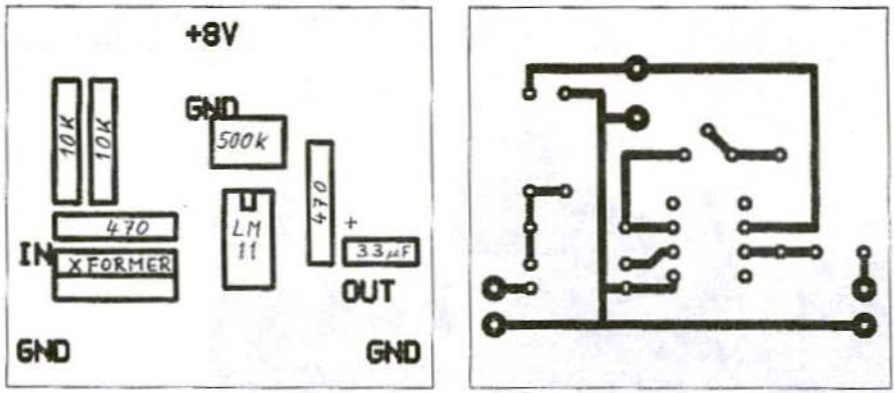

Fig.32c:Printed Circuit Board and Component Layout of the AF Amplifier

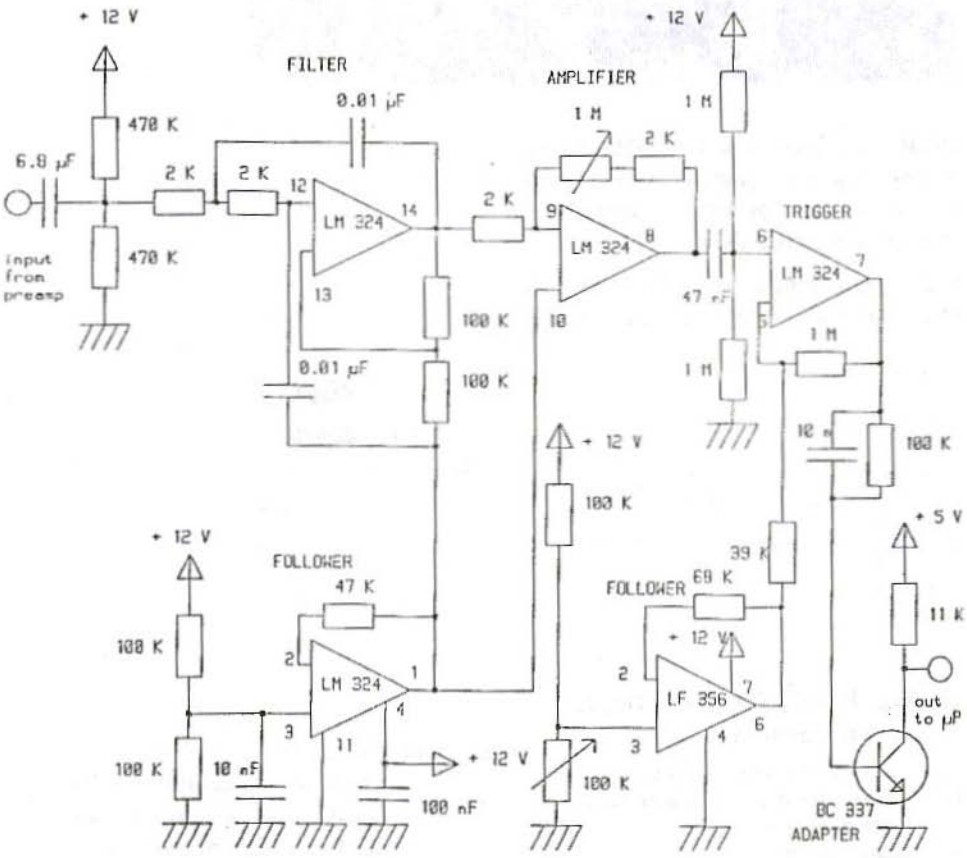

Fig.33a: Filter and Trigger Circuit

setup to measure distances from zero up to 500 metres. A lower pulse repetition frequency would probably give no additional range, even with a better 30 MHz IF unit barely more than 500 metres could be expected.

What we need to investigate now is how fast we can achieve frequency switching at 10 GHz. The 20V, 1 microsecond pulse for the varicap diode must have an internal resistance as low as possible in order that the pulse edges are as steep as possible (Fig.28).

We hope then that the 10 GHz oscillation (that's still 10,000 oscillations in one microsecond) follow on directly from the change in capacity, so that bursts as clean as possible are sent. During reception, when the signal is mixed down to 30 MHz, we shall only be counting 30 oscillations in 1 microsecond; 30 is not many.

The amplifier and filter circuit must be optimised with care if accurate results are to be achieved. We could spread out the frequency count over several periods, to increase the accuracy. The count result must be multiplied by two then, in order to compensate for the 1µs gaps between the 1µs bursts (Fig.29).

This measure would make the critical filter between IF preamp and HF receiver superfluous (Fig.30). The aperiodic amplifier with a necessary amplification of 100,000 to 1,000,000 would not be entirely uncritical, however. The frequency counter would count for 2 seconds and each time be preset by 30,000,000.

If we now measure the speed of a car travelling at 100km/h, it will have moved on 56 metres during the 2 seconds measuring time. It would be advisable then to select a shorter measurement window. In the deliberations that followed, an opportunity for this occurred.

Our setup can be simplified in two ways:

a) If the doppler frequency lies between

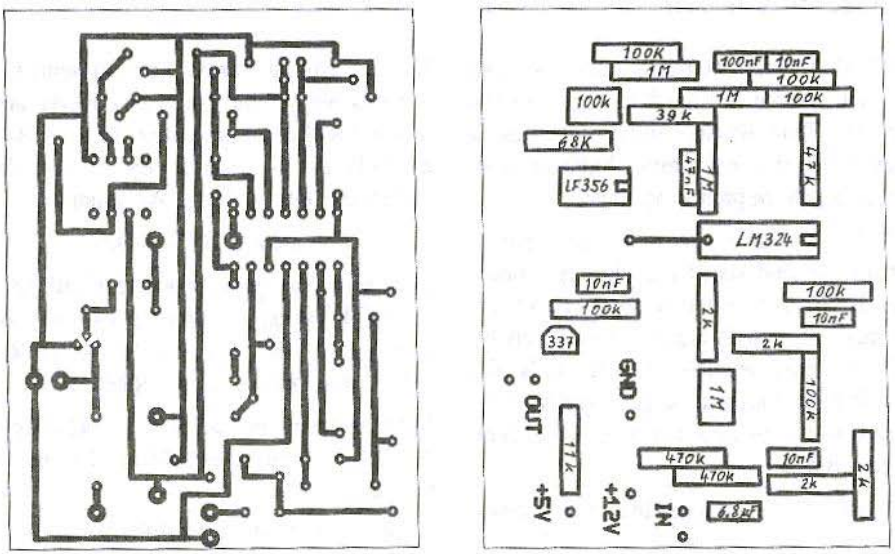

Fig.33b/c: Printed Circuit Board and Component Layout for Counter Display

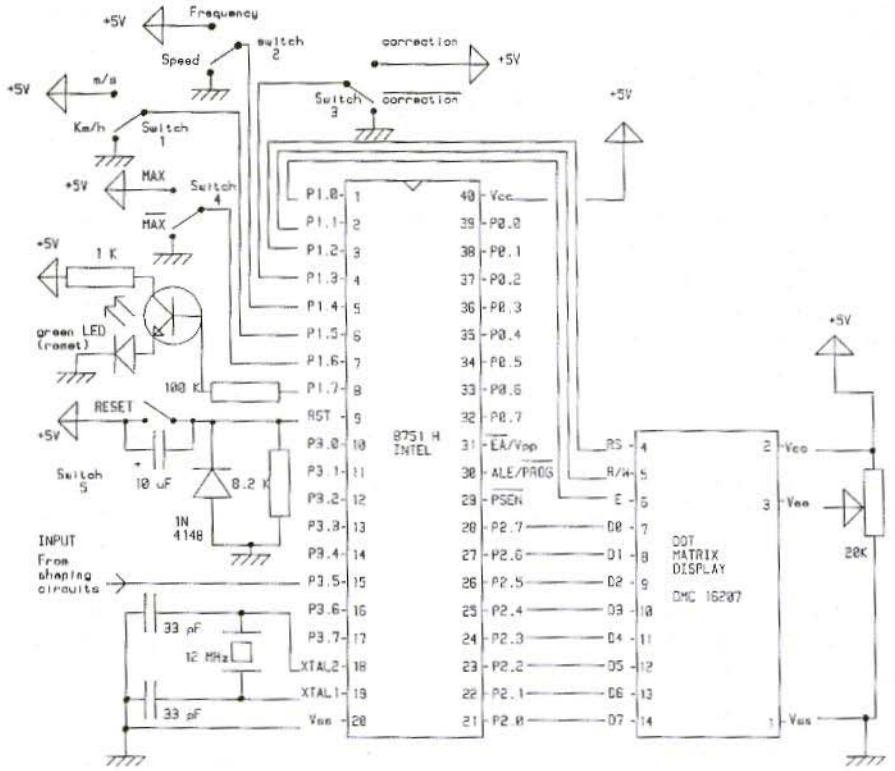

Fig.34a: Processor/Counter Display

30.000000 and 30.018000 GHz we can omit the three left-hand positions of the display. Our display, which only has to cover the five remaining digits, doesn't even have to be preset any more.

b) As we have seen that the doppler frequency and speed are directly proportional to each other, we can count the number of periods during an intentionally chosen time window and multiply the result by a constant, so as to produce the speed in the desired unit of measurement. The constant is:

$1.463 * 10^{-2}$ for a speed in metres/second

or:

$5.266 * 10^{-2}$ if km/h are desired.

If the counting period now measures 2 seconds we get the frequency in Hz and this gives the simplification that we can multiply this 2 seconds by one of the factors mentioned above. We count:

$$2s * 5.266 * 10^{-2} = 0.1053s$$

and the display gives us the result directly in km/h! During this short period the car driving at 100km/h has only moved 2.9 metres, and this is quite acceptable.

If we wanted the display in metres per second, the counter would need a gating time of:

$$2s * 1.463 * 10^{-2} = 0.029626 \text{ seconds.}$$

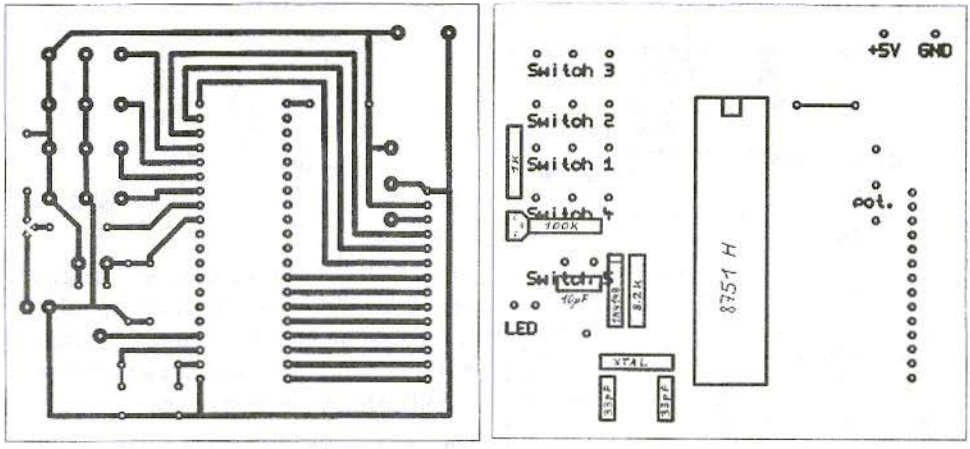

Fig.34b/c: PCB and Component Layout for Counter and Display

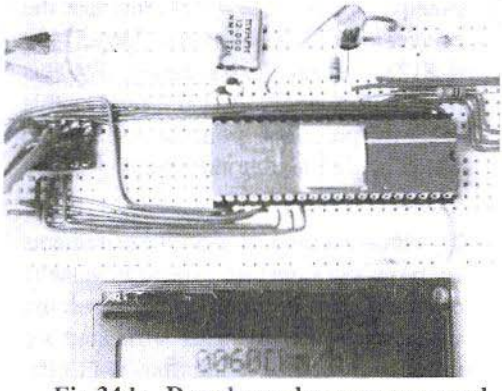

Fig.34d: Doppler radar processor and Display

4. CALCULATING SPEED

The dot-matrix LCD used (Fig.34a) needs to be driven by ASCII commands such as CLEAR, cursor position and others. For the calculating tasks we use an Intel 8751 microprocessor, which works with 8-bit words. With 8 bits we can count up to 255 (decimal), meaning that a doppler frequency of, say, 2000 Hz would be

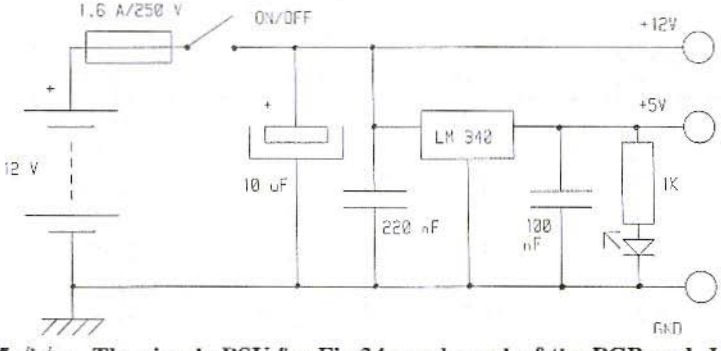

Fig.35a/b/c: The simple PSU for Fig.34a and overleaf the PCB and Layout

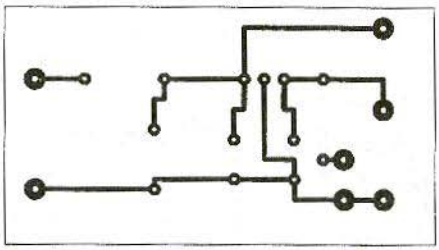

Fig.35b

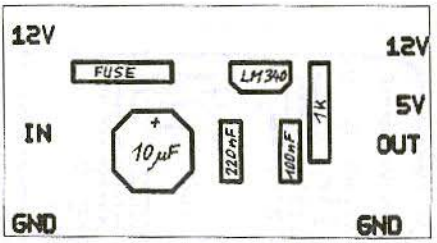

Fig35c

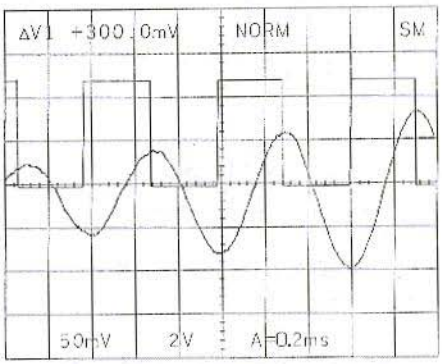

Fig.36: I/P and O/P signals of the Schmitt Trigger

too large. Fortunately, however, the timer embedded in the microprocessor uses 16-bit words.

The selected algorithm counts the incoming doppler pulses over a predetermined time span. With a Radar transmit frequency of 9.4 GHz (without doppler shift) the gate period is 15.947ms if we require speed in metres per second. If we use a different transmit frequency, the program must be modified for the new gate times.

The timer in the processor must be loaded with the complement of the gate time to 65536 ($=2^{16}$). That means, taking an example of gate time = 57409us, that the counter must be loaded with 65536 - 57409 = 8127. It then counts during 57409us from 8127 to 65536. When it reaches this value the processor produces an interrupt which is used to stop the counting of the doppler impulses.

To increase accuracy, eight measurements are taken and from these the mean value is calculated and displayed. The maximum values of all measurements carried out are stored and recalled if the time is insufficient for eight measurements. We assume that the highest value is the best and display this.

F input (Hz)	Display (km/h)	Calculated value (km/h)	Display (m/s)	Calculated value (m/s)
10	0	0.52	0	0.14
50	2	2.59	0	0.72
100	5	5.18	1	1.44
500	25	25.9	7	7.20
1000	51/52	51.8	14	14.40
5000	259	259.1	71/72	71.97
10000	518	518.2	143/144	143.95

Table 2

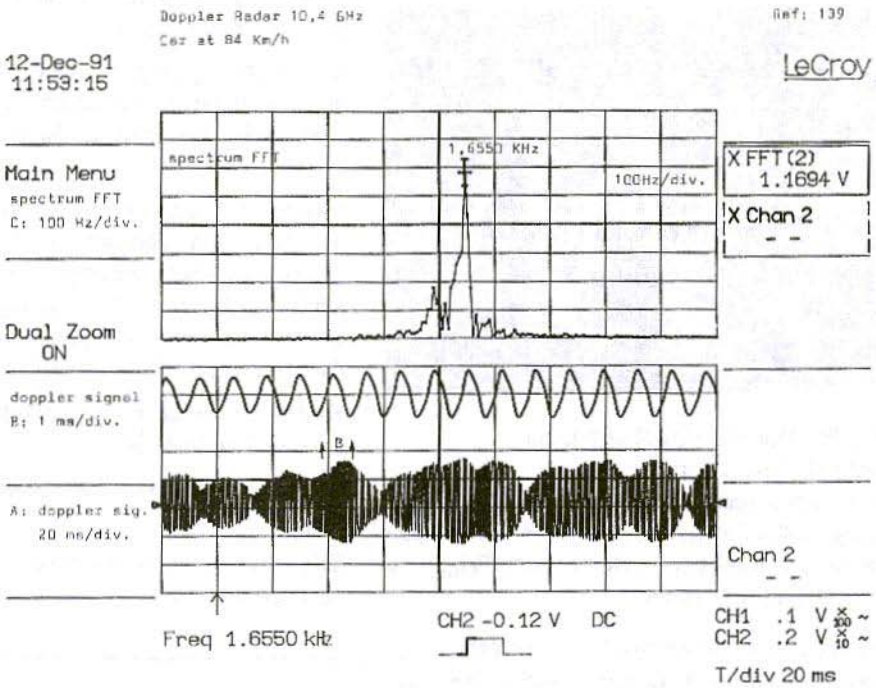

Fig.37: Lower curve shows Doppler signal of a car at 86km/h. Signal variations are clearly visible. Above this is represented extract B; the signal is deviating from a sine wave format. The upper display shows Fast Fourier Transform with a main frequency of 1.6550 kHz

5. PRACTICAL CONSTRUCTION

For our practical trials we wanted to use something simpler. An unkeyed blast-through mixer, very similar to the broad-band FM Gunn transceiver we already had, turned out to be good enough for a range of up to 100 metres.

Figures 31a to 31f show sufficient mechanical details for radio amateurs with some experience of these techniques to see what we have in mind.

The basis is a piece of R100 waveguide with internal measurements 22.86mm x 10.18mm. The Gunn element employed in the oscillator is a CXY11C. Adjoining this but separated by an iris window is the receive mixer with a 1N23C (Fig.31g).

The dish antenna used is by Sivers Lab and has a diameter of 36cm. This should give an antenna gain of 25dB at 9.5 GHz.

Fig's.32a to 32c show the entire electrical arrangements and the low-noise audio frequency preamp in circuit and layout. An LM11CN is used here.

Fig's.33a to 33c show the following stages: filter and amplifier, again in circuit and layout. After adequate amplification

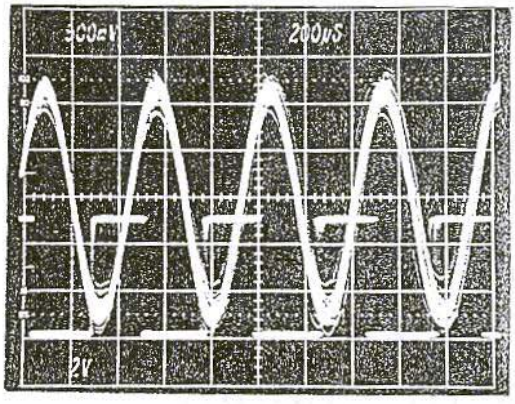

Fig.38: Doppler signal of a train 2030 Hz / 105,2 km/h
The high Radar cross-section value produces a steep (900mV/division) and low-noise Doppler signal; the pulses from the Schmitt Trigger are correspondingly clean

the signal is transformed into a rectangular form in a Schmitt Trigger. Finally this signal is passed to pin 15 of an Intel microprocessor which carries out the counting and calculation (Fig's.34a to 34d). The circuitry for the stabilised power supply is shown in Fig's.35a to 35c.

5.1 External connections

Five switches and one LED provide the user interface. They have the following operations:

Switch 1: toggling the display between km/h and m/s.

Switch 2: toggling between displaying doppler frequency and speed. When frequency is displayed switch 1 has no effect.

Switch 3: switches in a correction that takes into account the 33.6 degrees angle between the road and the Radar direction. If this is angle is neither zero or 33.6 degrees, then the software must be modified.

Switch 4: puts the highest value of the series on the display.

Switch 5: resets the microprocessor.

The LED is illuminated each time the micro is reset.

Frequency (Hz)	Amplitude (mV_{pp})
3200	1.4
7000	2.0
8400	3.0
11000	10
14000	100
20000	1000

Table 3

6. PRACTICAL TESTS

On the laboratory bench we were feeding in known frequencies directly into the low-frequency evaluation circuitry and obtained the speed results indicated in Table 2. As we can see, the accuracy of the counter is within plus or minus 1km/h.

6.1 Sensitivity

Table 3 shows the minimum input voltage needed at the preamplifier to produce a stable signal at the output of the Schmitt Trigger (Fig.36).

As can be seen, our direct conversion receiver is fairly deaf! Of course higher receiver sensitivity would increase the range of our Radar - this has been discussed already.

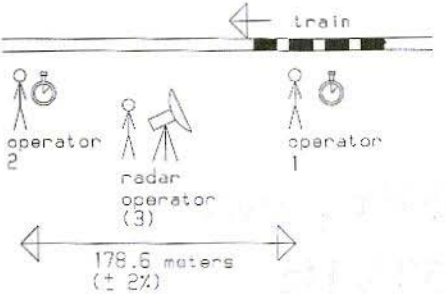

Fig.39: Free space test and Calibration setup

The screen shots (Fig's.37 and 38) show how the system works in practice. In addition the doppler signals produced by trains and cards are recorded at audio level on tape and later shown on an oscilloscope in the laboratory. It is particularly interesting to note here that real doppler signals are not as clean as the ones produced artificially in the lab, but on the other hand, they are not spurious readings.

Practical checking of accuracy and range was achieved as in Fig.39. When a train comes operator 1 signals the exact time the locomotive passes and operators 1 and 2 start their watches. When the locomotive passes operator 2, he gives a signal and they both stop their watches. The average of the two times is taken and operator three reads off the Radar display.

In three tests with varying speeds we found inaccuracies of from 4 to 6 per cent, the variations in the stopped time included. The maximum distance at which we could measure the train was around 100 metres. Possible improvements have already been discussed. If there is no Radar signal on the input, amplifier noise reaches the trigger stage and the display shows spurious readings. This spurious display could of course be suppressed with further software development.

To close we would like to offer cordial thanks to Pierre-Alain Glardon for his intensive and excellent co-operation in this project.

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